Dr. (Eng.) Ralph Oppelt, DB 2 NP

Stabilizing the VCO Frequency by Means of Monostables
Part 1

Expressions such as „monostable multivibrator" or "monoflop" have, since digital chips became cheap and ubiquitous, encroached into the mainly analogue world of HF circuit technology, e.g. (1) and (2). It is therefore appropriate to conduct a short overview of the principles of working and particular characteristics of such modules in their function as frequency stabilizers as well as – and without any claim to completeness – a series of application possibilities. A second article will concern itself with a constructional project of an SSB VCO for the 2-metre band.

1. PRINCIPLE

Fig. 1 shows the block diagram of a VCO stabilized with the aid of a monostable. The frequency range $f_L \ldots f_H$ of the VCO is translated by means of a stable reference frequency f_{xo}, derived from a crystal oscillator (XO), to a relatively low frequency. The output of the first low-pass filter LP_1 suppresses the sum frequencies $f_L + f_{xo} \ldots f_H + f_{xo}$ and passes the difference frequencies $f_L - f_{xo} \ldots f_H - f_{xo}$.

Instead of the "upright" mixing occasioned by f_{xo} being smaller than f_L, a small modification to the feedback loop allows an "inverted" mixing process whereby f_{xo} is greater than f_H. [The terms "upright" and "inverted" refer to the products of the modulation process, whether they occupy the same relative dispositions as the untranslated range of frequencies, low to high, i.e. upright, or run high to low; inverted. – G3ISB]

The following treatment will presume an upright mixing process. This results in the output of the limiter/amplifier containing a symmetrical rectangular pulse having a frequency spectrum:

$$f_{min} = f_L - f_{xo} \leq f \leq f_{max} = f_H - f_{xo} \qquad (1)$$

This triggers the following monoflop with either the positive or the negative pulse flank.

As shown in **figure 2**, the monoflop's output is a train of rectangular pulses $U_a(t)$ with a constant

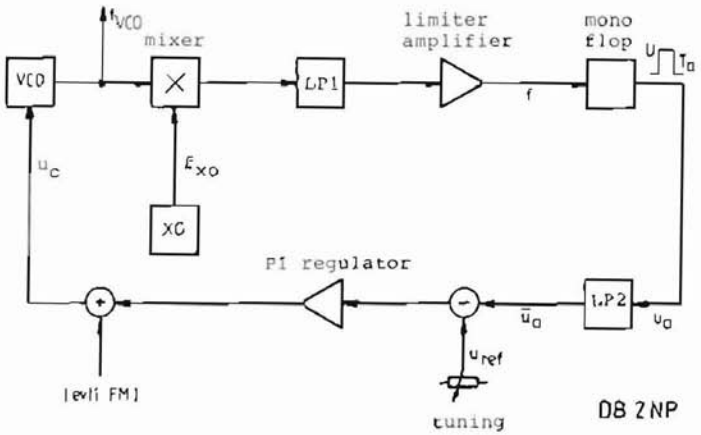


Fig. 1: Principle of frequency stabilization with a monostable

pulse duration of T_a and which have a varying pulse repetition frequency f in the range:
$f_{min} \leq f = 1/T \leq f_{max}$

The low-pass filter (LP 2) has a limit frequency $f_t << f_{min}$ therefore its output is $\bar{u}_a$, i.e. the integrated version of u_a (t) taken over time period T. With exactly rectangular signals, as depicted in figure 2, the average of u_a (t) is given simply by the ratio T_a/T

$$\bar{u}_a = U\,T_a/T = U\,T_a\,f \qquad (2)$$

In order that the monoflop is always reset to zero within the period T, the condition $T_a < T$ must be fulfilled for the total frequency range $f = 1/T$, or in other words:

$$T_a = k\,T_{min} = k/f_{max} \qquad (3)$$
with $0 < k < 1$

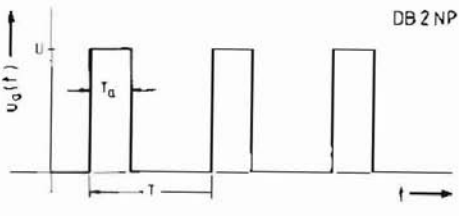


Fig. 2: Typical output signal of a monostable being triggered at a frequency f = 1/T

The following Performance Index regulator (PI) controls the VCO frequency f_{VCO} as long as $\bar{u}_a$ coincides with the (as stable as possible) reference voltage u_{ref}. From the equation (2) it follows for f_{VCO}:

$$f_{VCO} = f_{xo} + f = f_{xo} + \frac{\bar{u}_a}{UT_a} = f_{xo} + \frac{u_{ref}}{UT_a} \qquad (4)$$

As f is directly proportional to u_{ref}, the VCO frequency can be varied linearly by a linear potentiometer controlling u_{ref}.

The practical advantage of this characteristic is self evident. The potentiometer can be directly calibrated in terms of frequency. This can be arranged easily in a transceiver, for example, as the indicator has only to deal with DC instead of RF. Preset channels (e.g. transponder frequencies) can also be arranged easily by means of fixed potentiometers as well as the construction of sweep-oscillators. The latter manual tuning application requires that the tuning voltage be applied in the form of a saw-tooth or a triangular wave.

As fig. 1 shows, frequency modulation is easily carried out by the application of AF at this point. Finally, it should be mentioned, that a control loop containing monoflops, as those with PLL chips, are also suitable for the demodulation of FM.

2. THE STABILIZING EFFECT

Given the pre-condition that the monoflop is able to work with the high VCO frequency, the control concept of the downwards mixing process presented in fig. 1 is not, in principle, required. Because a relative VCO frequency error $\Delta f/f_{VCO}$ is converted at the input to the monoflop, into the considerably larger frequency error of $\Delta f/f$, the frequency stability is noticeably increased (f_{xo} can be regarded as being absolutely stable). The error sensitivity increases by a factor:

$$\frac{\Delta f/f}{\Delta f/f_{VCO}} = \frac{f_{VCO}}{f} = 1 + \frac{f_{xo}}{f} = 1 + \frac{f_{xo}}{f_{VCO} - f_{xo}} \qquad (5)$$

The frequency shift $f_H - f_L$ of the VCO is normally very much smaller than the lowest VCO output frequency f_L. For example, the VCO shifting over the 2 m band: 2 MHz << 144 MHz. According to equation (5), f_{xo} must be disposed closely below f_{VCO} in order to obtain greater error sensitivity. The main problem now arises, that there is insufficient decoupling between the two mixer inputs resulting in frequency f_{xo} appearing in the VCO output signal, which should, in principle, be free of any other spurious frequencies. Also, because of the proximity of the f_{xo} to the output frequency f_{VCO}, the following buffer stage is not able to assist in the suppression of f_{xo}. This disadvantage can be clearly seen in (1).

Some assistance may be afforded here by the use of an isolating amplifier in the form of a directional coupler between the VCO and mixer, or – easier to implement at UHF – a double-downward translation to bring the frequency of the first XO to a greater spectral distance from f_{VCO}.

In some circuit examples, the downward translation is replaced by a divider chain. This is only worth-while when the VCO itself already covers a large frequency range ($f_H/f_L \geq 2$) as shown, for example in (2). As the error frequency Δf of the VCO is also divided down by the same division factor as the signal f_{VCO}, the sensitivity gain given in equation (5) is no longer valid.

2.1. The Effect of Interference

By analyzing the influence of spurious signals, a few general observations may be made.

Assuming that a disturbing quantity is a noise voltage which is superimposed upon the VCO control voltage u_c. The effect of the noise upon u_c and $\hat{u}_a$ will be smaller as the deviation is increased. A large deviation of u_c means that the control sensitivity of the VCO should not be made too great (usually designated as [K_f], which is the same as Hz/V).

The VCO control voltage u_c is usually applied to a varicap diode. With a portable equipment using a battery voltage of 12 V, a 3 to 8 Volt variation should be necessary to cover the bandwidth f_L to f_H. A larger deviation for $\hat{u}_a$ demands, first of all, the largest possible value for T_a, i.e. the highest value for k in eq. (3). T_a/T_{min} is then only just short of unity.

Secondly, T_a/T_{max} must approach zero as nearly as possible. This can be accomplished by the correct choice of f_{xo}. The voltage deviation range $\hat{u}_{a\,max} - \hat{u}_{a\,min}$ can be obtained from eq. (2) and (3).

$$\hat{u}_{a,max} - \hat{u}_{a,min} = \frac{Uk}{f_{max}} \cdot (f_{max} - f_{min})$$

$$= Uk \left(1 - \frac{f_{min}}{f_{max}}\right) \qquad (6)$$

Given the VCO boundary frequencies f_L and f_H the required monoflop pulsewidth T_a and the oscillator frequency f_{xo} can be determined.

$$T_a = \frac{\hat{u}_{a\,max} - \hat{u}_{a\,min}}{U\ (f_H - f_L)} \qquad (7)$$

$$f_{xo} = \frac{f_L \cdot \hat{u}_{a\,max} - f_H \cdot \hat{u}_{a\,min}}{\hat{u}_{a\,max} - \hat{u}_{a\,min}} \qquad (8)$$

The maximum obtainable deviation $0 \leq \hat{u}_a \leq U$ should only be utilized to 80 %. For example,

$u_{a\,min} = 0.1\ U$ and $u_{a\,max} = 0.9\ U$ could be chosen. For these practical values inserted in eqs. (7) and (8):

$$T_a = \frac{0.8}{f_H - f_L} \qquad (9)$$

$$f_{xo} = (9\,f_L - f_H)/8 \qquad (10)$$

Of course, the calculated value for f_{xo} need not be strictly adhered to, but it does serve as a working basis. A suitably cheap crystal which comes closest to that calculated above is used.

2.2. The Effect of Drift

Following these more basic dimensioning considerations, the effects of disturbing influences – above all drift – will be analyzed. The most important factor for stability is the reproducibility of:

a) the monoflop pulsewidth T_a
b) the Voltage U
c) the reference voltage u_{ref}

Short-term instabilities, in the form of needle impulses, can be attributed to VCO phase-noise. This type of variation is manifest as phase-jitter in (a) and superimposed noise voltage in (b) and (c). Low phase-noise is ensured by low-noise voltages U and u_{ref} as well as the provision of a low-noise operational amplifier for the PI regulator.

Contrary to the practice in sweep oscillators, the loop control speed of the manually tuned VCO plays a relatively unimportant role. This allows a low-pass filter to be placed behind the PI regulator which allows only the low-frequency noise components through and thereby concentrating the VCO noise pedestal directly about the VCO output frequency.

Disturbing elements also take the form of the above discussed quantities but they are long-term and are caused by thermal drifts. As the regulator is forced continually to make $\bar{u}_a = u_{ref}$, a variation of the pulsewidth T_a, to $T_a' = T_a + \Delta T_a$ eq. 2 for the frequency error Δf is:

$$\bar{u}_a/U = u_{ref}/U = T_a f = (T_a + \Delta T_a)(f + \Delta f) \qquad (11)$$

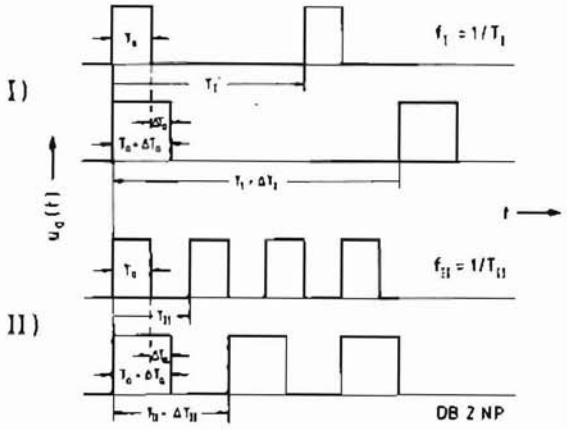


Fig. 3: Frequency drift as the consequence of a – greatly exaggerated – drift in the monostable pulsewidth T_a to $T_a + \Delta T_a$, by a factor 2.5 for the two monostable frequencies f_I and f_{II}. As the PI regulator always retains the same average value $\bar{u}_a$ (here U/5 or U/2), the duty factor remains constant: $T_a/T = (T_a + \Delta T_a)/(T + \Delta T)$. With $f = 1/T$ and $f + \Delta f = 1/(T + \Delta T)$ results directly in eq. 12. This means that with any given drift ΔT_a, that $\Delta f/f$ remains constant, i.e. a constant *relative* frequency error. With $f_H = 2.5\ f_L$, it follows that $\Delta f_{II} = 2.5\ \Delta f_I$

$$\Delta f = -f \frac{\Delta T_a}{T_a + \Delta T_a} \qquad (12)$$

The VCO frequency is lowered when T_a increases and, interestingly enough, more sharply the higher the monoflop's input frequency f was before the T_a drift took place. It follows, that from the previously mentioned reasons for f to have a large deviation i e $f_{max}/f_{min} >> 1$, that the VCO's stability is noticeably better at the lower than at the upper band-limit

Fig. 3 demonstrates these effects in diagrammatic form ΔT_a is therefore made disproportionately large in order to show the effects clearly. Working through this example shows the proportions between Δf and f as found in eq. 12.

Next, the effect of the drift in the monoflops supply voltage U is considered as it changes to U' = U + ΔU. Since U in eq. 2 may be treated in exactly the same way as T_a, one receives directly by substitution of U and T_a in eq. 11

$$\hat{u}_a/T_a = u_{ref}/T_a = Uf = (U + \Delta U)(f + \Delta f) \quad (13)$$

The solution for Δf yields again eq 12:

$$\Delta f = -f \frac{\Delta U}{U + \Delta U} \qquad (14)$$

This may be seen in the numerical example of fig. 4

The effects of both, a drift in U and in T_a, lead to a higher frequency drift at the VCO's higher frequency f_H than at its lower frequency f_L. If the frequency stability at the upper-band end is not satisfactory, the range $f_L...f_H$ can be divided into n equal band segments in order to make it better by a factor of n. The n switchable crystal frequencies would represent an additional complication.

Now, the drift in reference voltage u_{ref} is considered, $u'_{ref} = u_{ref} + \Delta u_{ref}$. This analysis of the attendant frequency error Δf is quite simple as f is directly proportional to u_{ref} (as it should be), so eq. 4 follows directly:

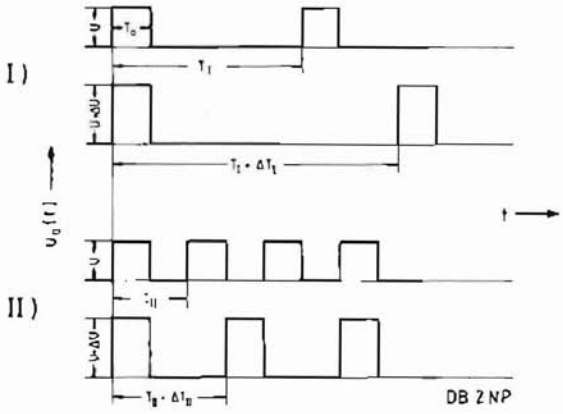


Fig. 4: Frequency drift as the consequence of a – greatly exaggerated – drift of the monostables output amplitude from U to U + ΔU, by a factor of 2.5, of the two monostable frequencies f_I and f_{II}. Here, also, the PI regulator forces the average value $\hat{u}_a$ to be maintained (U/5 at f_I and U/2 at f_{II}). As, this time, T_a is an assumed constant, it follows that the same amplitude/period ratio prevails prior to, and following the drift: U/T = (U + ΔU)/(T + ΔT). With f = 1/T and f + Δf = 1/(T + ΔT) results directly in eq. 14.

$$\Delta f = \frac{\Delta u_{ref}}{U T_a} \qquad (15)$$

A drift of u_{ref} has the same effect in all parts of the VCO's tunable range $f_L \ldots f_H$.

It is now conceivable that the frequency drift component, caused by a change in temperature, can be countered by superimposing a temperature-compensation voltage on to u_{ref}. This could be done with an NTC-resistor network, for example. This compensatory voltage would correct the VCO frequency by the same amount, higher or lower, and independent of frequency f according to eq 15.

Frequency drifts caused by ΔT_a or ΔU have to be corrected additionally, with f taken into account. See eq. 12 and 14. As, nowadays, supply volatges U and reference voltage u_{ref} may be made very stable and reproducible owing to the availability of high-grade voltage-regulator chips, the main cause of frequency drifting is due to the drift in the monoflop's pulse-width T_a. This is countered normally by making an external RC network, a small part of which contains an NTC or PTC resistive element. This offers a simple means of temperature compensation of the T_a drift.

3. THE USE OF MONOFLOPS IN FM-FIXED FREQUENCY OSCILLATORS

Fixed-frequency-channel, AM equipments using crystals, employ transmitter concepts involving the RF output signal being derived from the receiver L.O. (e.g. 135 MHz) being mixed with the IF frequency (10.7 MHz). This device enables the receiver crystal (about 45 MHz, working in 3rd over-tone) to be used for the transmitter, thus saving both space and money.

Operation with FM, however, requires that the crystal be a relatively low frequency and therefore must be multiplied many times to achieve 135 MHz entailing the use of inter-multiplier filtering. This multiplication process ensures that a sufficiently large system deviation is obtained (which normally requires a multiplication factor of 9 or 12). As before, the frequency channelling requires a large number of FM crystal oscillators to be switched and all multiplied to the same frequency band – not an easy matter.

Commercial hand-held transceivers using fixed (crystal controlled) channels usually employ unmodulated crystal oscillators. A phase modulator stage follows the crystal oscillator and the AF input is taken through a I/F network. The resultant modulated RF output then simulates an FM waveform which has been generated in a more conventional manner. This method also relies heavily upon frequency multiplication (at least x 9) in order to achieve the required deviation. The Oberon project (ref. 3) solved this problem by using a complete PLL with a fast pre-scaler. This is actually more by way of being a frequency synthesizer as opposed to the discrete transmit and receive crystals considered here.

There is a concept which carries out the modulation process at IF. This uses a monoflop stabilized, fixed-frequency oscillator and is capable of easily generating the required deviation without recourse to multiplication according to the principle of fig. 1. For relay work, however, the IF must be able to be switched to 10.1 MHz from the nominal 10.7 MHz IF. If the crystal oscillator XO of fig. 1 is required to have only one crystal frequency (e.g. f_{xo} = 10 MHz), the direct channel (f = 0.8 MHz) will experience noticeably greater drift problems than that of the radio relay channel (f = 0.1 MHz) (see eq. 12 and 14). This may be optimized by choosing a middle value for f_{xo} from both IFs, namely 10.4 MHz.

The use of 10.1 MHz involves a downward translation process resulting in an "inverted" output as $f_{xo} > f_{VCO}$. In order that the control loop works in a negative feedback manner, this case must be provided with a pre-inversion.

This can best be done by switching from the monoflop's output u_a to its complementary output u'_a. A further switch-over process is not necessary as the monoflop frequency is f = 300 kHz in both cases when f_{xo} = 10.4 MHz. This means, of course, that u_{ref} does not have to be changed.

A crystal with an awkward frequency of 10.4 MHz can prove difficult to obtain and might be expensive. Considerably easier to obtain would be 10.245 mixer crystals and 10.345 MHz PLL crystals which also lie more or less between 10.1 and 10.7 MHz. Their use, also, does not entail the necessity of switching u_{ref}, if T_a and u_{ref} are suitably chosen:

Assuming a chosen crystal frequency f_n between f_L = 10.1 MHz and f_H = 10.7 MHz, the monoflop frequencies are:

$f_1 = f_H - f_{xo}$ (upright-downward mixing) (16a)

$f_2 = f_{xo} - f_L$ (inverted-downward mixing) (16b)

From this, the average value of the monoflop output voltage results:–

$$\bar{u}_{a1} = U T_a (f_H - f_{xo}) \quad (17a)$$

$$\bar{u}'_{a2} = U [1 - T_a (f_{xo} - f_L)] \quad (17b)$$

It should be noted that in eq. 17b, the inverted output of the monoflop is selected.

If the change-over of u_{ref} is to be avoided, both voltages must be equal in amplitude i.e.:

$$u_{ref} = \bar{u}_{a1} = \bar{u}'_{a2} \quad (18)$$

From eq. 17a and 17b it follows that:

$$T_a = \frac{1}{f_H - f_L} \quad (19)$$

$$u_{ref} = U \frac{f_H - f_{xo}}{f_H - f_L} = \frac{U}{T_a} (f_H - f_{xo}) \quad (20)$$

The monostable pulsewidth T_a is, interestingly, not only independent of the crystal frequency f_{xo} in use, but also fixed upon the reverse value of the relay i.e. 1/(600 kHz).

The method by which a frequency modulated 10.1/10.7 MHz oscillator can be realized is demonstrated by the example of fig. 5. This circuit sketch is a little less than a constructional project but it does serve to show that by using cheap consumer ICs, the component count and therefore the constructional size, can be kept small. With a little PCB-design experience, the whole arrangement of fig. 5 can be placed on a board of only 4 x 7 cm².

The AM receiver chip Siemens TCA 440 functions here as a downwards mixer, crystal oscillator, limiter amplifier and buffer between VCO and mixer – this buffering is to suppress the f_{xo} spurious frequencies, see chapter 2.

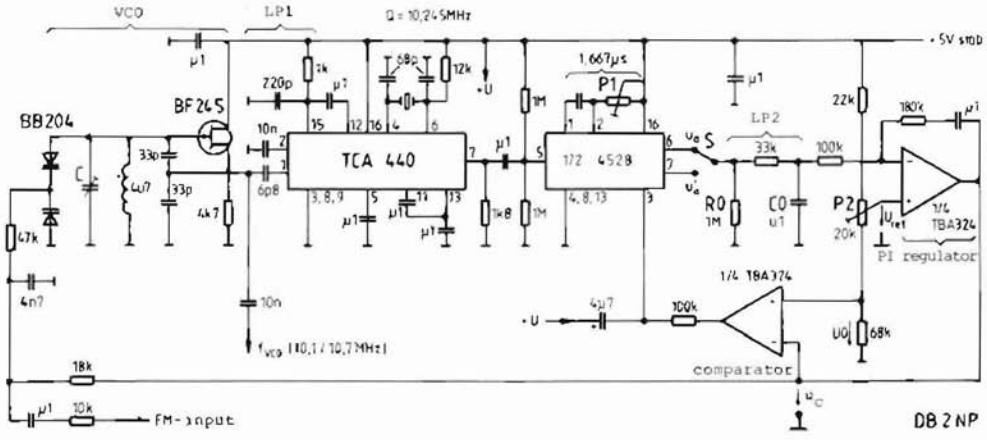


Fig. 5: Circuit diagram of a sample, frequency-modulated 10.1/10.7 MHz oscillator.

The IF. square-wave signal at pin 7 has about 2 V_{pp} amplitude and is sufficient to trigger the CMOS monostable 4528, the input of which has been biased at U/2.

The pulsewidth T_a at pin 6 of the monostable is adjusted to 1.667 μs with the aid of P1 (see eq. 19). Potentiometer P2 adjusts the reference voltage u_{ref} of the PI regulator in accordance with the crystal frequency f_{xo} = 10.245 MHz (see eq. 20). P1 and P2 are then fine trimmed with f_{VCO} being monitored on a frequency counter. When switch S is in the u_a position, the f_{VCO} = 10 7 MHz exactly. Likewise with S in position u'_a, f_{VCO} = 10.1 MHz.

In order that the VCO does not jump out of lock during the switching of S, a comparator has been added to the circuit. Immediately following every switching action, the previous VCO frequency is always on the wrong sideband so that the loop has a positive feedback This means that the initial drift of the control voltage U_c – irrespective of direction – is amplified At the moment of switching, switch S has neither contact with u_a nor with u'_a. The voltage at C_o sinks momentarily due to R_o thus causes the control voltage U_c to drift upwards.

When switching the lower to the upper frequency, the favourable direction is: The VCO frequency gets to the correct sideband so that the VCO locks firmly on to 10.7 MHz. Switching back to 10.1 MHz, the VCO would run to its upper frequency limit as, this time, the control-voltage drift during the switching action is going in the wrong direction Because of the comparator, a further triggering of the monostable is inhibited when the control voltage exceeds U_o. The monoflops output u'_a assumes the voltage U. As a consequence, the control voltage U_C sinks quickly, and the VCO locks to the frequency of 10.1 MHz following the resetting of the comparator The trimmer C in the VCO tuned circuit must be adjusted such that the control voltage U_c at f_{VCO} = 10 7 MHz and is about 20 % below the value of U_o.

In conclusion, for the individualist, it is mentioned that in the construction of a complete FM transceiver, the receiver forms part of the transmit-frequency locked loop in order to improve transmit-frequency stability. The function of the monostable can be replaced by a normal FM demodulator or a coincidence detector if a fixed IF is employed. A good temperature stability is afforded by the FM-receiver chip TCA 420 with a push-pull output which facilitates any temperature drift to be compensated by means of a subtracter stage The more popular S 041 and TBA 120 are not recommended for this purpose as they only possess a multiplier output.

(To be continued)

4. REFERENCES

(1) M Volkert, DF 4 SQ:
"Zennor – ein Miniatur-CW-Transceiver für 14 MHz"
cq-DL 5/1987, S 295 - 301

(2) J. Jirmann, DB 1 NV:
"A Spectrum Analyser for the Radio Amateur Part 2 – Concluding"
VHF COMMUNICATIONS 19 (1987), Ed. 4, P. 232 - 242

(3) H. Helpert, DJ 9 HH:
"Oberon 2 m-Sender-/Empfänger-Konzept zum Selbstbau"
cq-DL 5/1987, S. 288 - 292 und
cq-DL 6/1987, S. 354 - 359

Dr. (Eng.) Ralph Oppelt, DB 2 NP

Stabilizing the VCO Frequency by Means of Monostables
Part 2: A Continuously Tunable VCO for the 2 m SSB Band

Part 1 of this article described the principles of the stabilization of VCOs by means of monostable multi-vibrators and also the frequency drift behavior of these oscillators. This second, concluding article, will describe the constructional details of such a VCO which may be employed, for example, with the phasing method of SSB generation described in ref. (1). This would then form the basis for an economical home-constructed, SSB, 2-metre transceiver.

5. CHOICE OF MONOSTABLE

The most convenient integrated monostables are the 4047, 4528 and 4538 from the CMOS 4000 series, also the 74121, 74221, 74122 and 74123 together with their improved, pin-compatible equivalents, the 74422 and 74423 from the 7400 series. The latter series have also meanwhile sprouted a multiplicity of various types in the 74SN..., 74LS..., 74ALS..., 74S..., 74F..., 74HC..., 74HCT...etc. families.

As the ICs are mostly used in battery-supplied equipment, the ones having the lowest power consumption i.e. CMOS or 74HC types, are to be preferred. One thing to be watched, however, is that the minimum duration pulse (T_a) required can be generated in the proposed type to be employed. According to eq. 9 (part 1) the required T_a, for a favourable dimensioning, is about 20 % smaller than the VCOs reciprocal deviation $1/(f_H - f_L)$. For the 2-metre band then, the output pulse duration T_a is about 400 ns – this is able to be reproduced with certainty.

Every real square-wave pulse has, however, finite rise and fall times and mostly has an overshoot or undershoot as well. As **fig. 6** makes clear, it is always important that the monostable output signal u_a (t) has returned to zero before the next trigger pulse. This ensures the linear relationship, indicated in eq. 2, between the monostable frequency $f = 1/T$ and the average value $\hat{u}_a$ of the monostable output signal u_a (t) is valid for any output wave shape. Any departure from the ideal wave shape in u_a (t) will always result in a modification to the relationship given in eq. 2 between $\hat{u}_a$ and f. If u_a (t) has not decayed to zero before the next trigger sequence occurs, a linearity error at the high end of the VCO's band will be the result. It is important that the minimum pulse attainable should be regarded as that occurring from the zero at the leading edge of the pulse, to the instant the wave-form has reached a steady state zero at the trailing edge of the

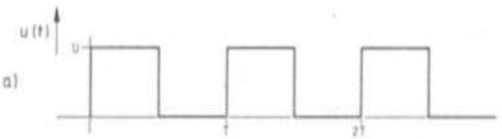


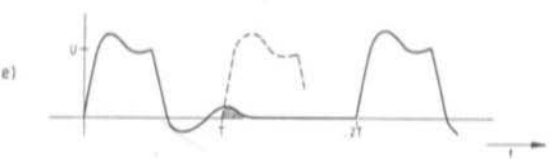
Fig. 6:
Influence of a non-ideal wave shape of a square-wave signal upon its average value $\hat{u}$, i.e. upon the actual frequency of the VCO

a) ideal square-wave, average value $\hat{u}_a = \frac{1}{T}\int_0^T u_a(t)\,dt = U/2$

b) as a) above but non-ideal square-wave, average value $\hat{u}_b = \frac{1}{T}\int_0^T u_b(t)\,dt = c \cdot U/2$ with $c \approx 1$

c) as a) but at half the trigger freq., average value $\hat{u}_c = (U/2 + 0)/2 = U/4 = \hat{u}_a/2$, i.e. exactly half the average value of a)

d) as b) but at half the trigger freq., average value $\hat{u}_d = (\hat{u}_b + 0)/2 = \hat{u}_b/2$, i.e. exactly half the average value of b)



e) Non-ideal square-wave with pronounced overshoot. The trigger frequency $\frac{1}{2T}$ average value

$\hat{u}_e = \frac{1}{2T}\int_0^{2T} u_e(t)\,dt$ (the solid curve is $u_e(t)$)

Triggering at twice the frequency $\frac{1}{T}$ (dotted curve), does not result in twice the average value $2\,\hat{u}_e$, as $\int_0^T u_e(t)\,dt + \int_0^{2T} u_e(t)\,dt$: $u_e(t)$ has, at time T, not reached a stable zero value (see also the hatched part).

pulse. This is the condition which the minimum attainable pulse duration T = 1/f will drive the VCO without any risk of linearity error.

The integrated circuit chosen for this 500 kHz tuning range, SSB VCO is the economical and easily obtainable 4528, in CMOS. This applies also to any VCO which is required to cover the whole of the 2-metre band i.e. 2 MHz. Nevertheless, there is quite a difference in these devices between those of different manufacturers. A good example of the 4528 should be able to produce a minimum pulse of about 120 ns duration with 5 V supply potential and an external load of $R_{min.}$ of 5 kΩ and output capacity of 0 pF.

6. VCO MODULE, CIRCUIT DIAGRAM

The circuit shown in the block diagram of **fig. 7** shows, besides the actual VCO, a crystal oscillator (XO) employed for the generation of a 9 or 10.7 MHz carrier for the mixer which translates the VCO frequency to the transmitting range of 144 to 144.5 MHz. This module may be combined directly with the SSB phasing-method generator described in (1) thus making for a compact single module.

In order that the problem with spurious signal generation, which was discussed in part 1, is avoided, the VCO uses two downward translations (despite what is shown in fig. 1) in order to attain the final local oscillator frequency. In order that things are kept as simple as possible, self-oscillating crystal mixers are used for this operation. An IF of 10.7 MHz was used which means that the VCO must cover from f_L = 133.3 MHz to f_H = 133.8 MHz for transmission on the lower quarter of the 2 m band. Both crystal mixers use cheap computer crystals, 24.576 MHz excited on its 5th over-tone giving an output on 122.96 MHz for the first one, and the second mixer uses a 10.24 or 10.245 MHz fundamental crystal oscillator.

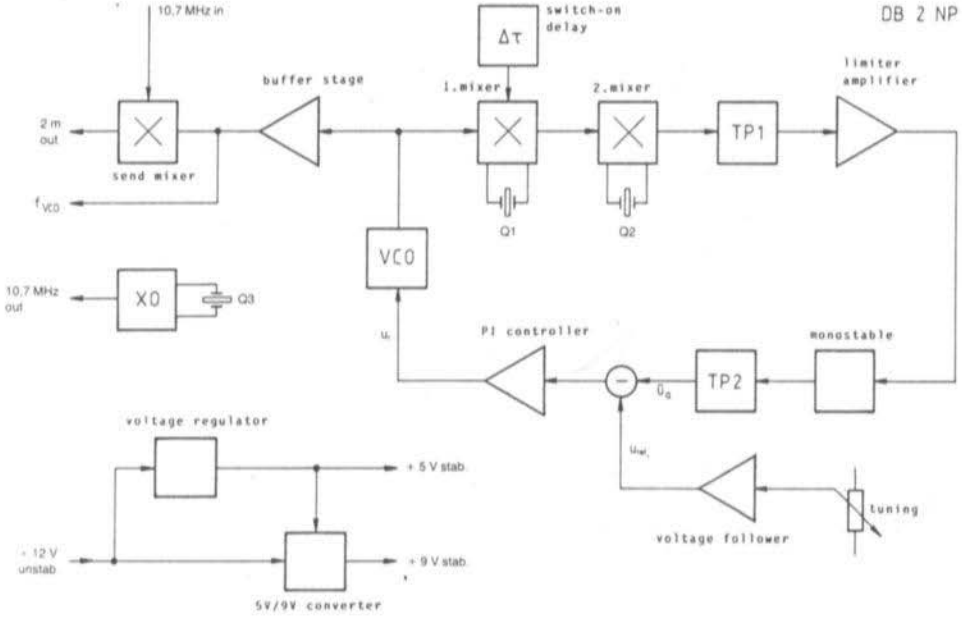


Fig. 7: Block diagram of the VCO for SSB operation in the 2 m band

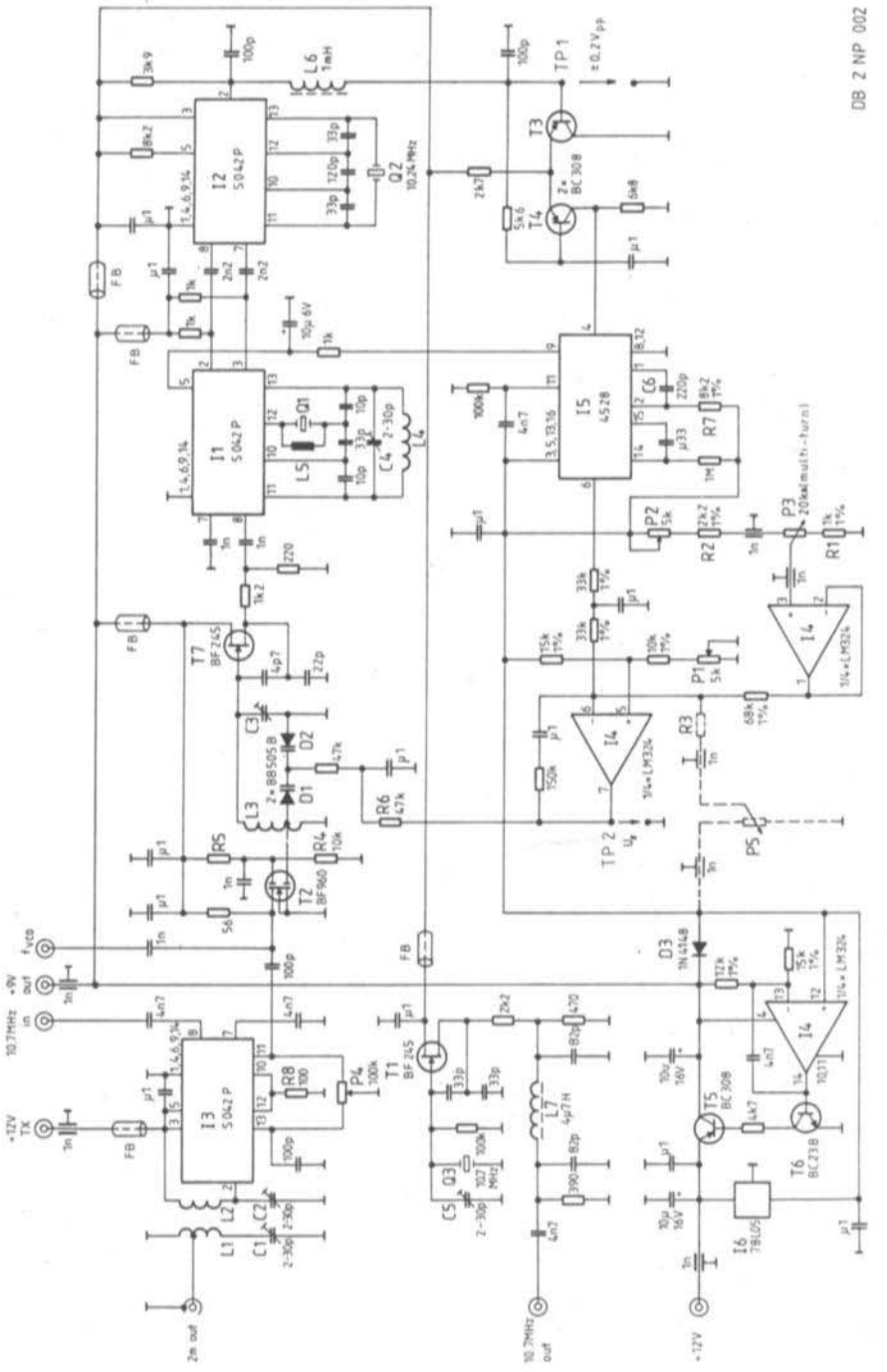


Fig. 8: The circuit schematic of the VCO of fig. 7

The circuit schematic of **fig. 8** shows the whole arrangement in detail. As many standard circuits have been employed, only the peculiarities in this design will be enlarged upon.

It may be observed that the varicap diodes D1 and D2 are not shunted across the complete tuning inductance L3 but only a low-impedance tapping point of it. This measure ensures that a larger tuning voltage increment is required to achieve a given frequency variation. Therefore, a reasonable tuning voltage range of 3 to 7 volts is required to cover the entire VCO range instead of a few millivolts (see part 1, para. 2).

The VCO signal is fed from the source circuit into the PLL control loop. It is then taken to a limiter-preamplifier (2 x BC 308) via the two mixer stages and a 3rd-order, 750 kHz, low-pass filter. It is then presented to the input of the monostable trigger. With the crystals used here, this trigger signal lies in the range of about 100 to 600 kHz. Using other combinations of Q1 and Q2 could result in a higher than intended spectrum resulting in a deterioration of the frequency stability.

The second monostable of the 4528 is disabled for a few milli-seconds following the application of the 12 V supply voltage. This ensures that the control loop trigger signal for the monostable is missing which simulates a "too low" VCO frequency. The control loop voltage U_c and with it the freely oscillating VCO, assumes its maximum possible value. Thereby, when the first mixer is activated via pin 5, the VCO is started up with the correct sideband and a proper PLL lock is ensured.

Frequency tuning is carried out by means of the 10-turn potentiometer P3. Owing to the somewhat coarse winding of this pot'meter, the tuning appears to occur in fine steps making it difficult to adjust to very fine limits. This effect is not pronounced enough to notice when tuning in a station. Should a more continuous tuning be required, it will be necessary to add the dotted-line components P5 and R3 and, depending upon the choice of R3 (MΩ range), it is possible to tune right down to ± 1 kHz or ± 5 kHz. Since the 10-turn pot'meter has a tuning rate of 50 kHz per turn, this alternative is, at the same time, a valuable tuning aid particularly for those whose hand is not too steady.

Of particular importance for the linearity is the resistor R1 on the ground side of P3 as it limits the voltage to be handled by the following voltage-follower IC to a few mV above its negative supply rail (i.e. 0 volts!). As the DC component of the monostable output voltage $\ddot{u}_a$ increases with the VCO frequency, a falling voltage must be provided at the output of the voltage-follower in order that the PI controller has a balanced voltage at its two inputs. When the slider of P3 is set against the connection of R1, the VCO frequency must be at exactly f_H = 133.8 MHz. This adjustment can be precisely adjusted with P1 at the non-inverting input of the PI controller. The setting of P2 is almost irrelevant as it does not affect the voltage set by P3 to any great extent as long as the wiper of P3, as ordained, is turned to the extreme end of its travel to the R1 end. If the wiper is now turned to its other extreme i.e. to the R2 end, the lower band frequency of f_L = 133.3 MHz may be set without affecting the upper frequency setting. This measure avoids the tiresome, iterative adjustments at lower and upper frequency limits so prevalent with other VCO designs.

The stabilized voltage for the monostable IC is provided by the standard three-terminal 78L05. It has proved fully satisfactory for this application but can, of course, be changed for a higher-quality type if required. This stabilized 5 V supply is also used, together with a second IC, type LM324, to provide a stabilized 9 volt for most of the other stages as well as for eventual external employment. Because a pnp type was chosen as the series stabilizer transistor only a 0.3 V voltage difference was necessary. It functions thereby perfectly satisfactory even when the battery voltage has sunk from 12 V to as low as 9.3 V (that is, in any case, long after a change of batteries should have been considered).

The function of diode D3 is of importance. This lies, in the same 9 V stabilized line as the LM324 and ensures that the latter is initially supplied with 4.3 V (i.e. 5 V minus the diode drop) upon switch-on. The circuit is then able to rise automatically to the required 9 V. Diode D3 is then poled in the blocked direction.

7. CONSTRUCTION, COMMISSIONING AND ADJUSTMENTS

Fig. 9 shows the component layout plan of the 72 x 109 mm PCB DB 2 NP 002 which is intended for a tin-plate box of 74 x 111 x 30 mm dimensions. The construction follows closely the procedure mentioned in (1); i.e. first drill all the holes in the tin-plate box, then solder the pre-drilled PCB into the housing – the ground plane to the walls of the box. The tin-plate compartment walls are then fitted and soldered in. The latter are, in fact, not strictly necessary but at least the VCO should be soundly enclosed and fitted with its own lid. The latter should also have a hole drilled in the appropriate place to allow C3 to be adjusted (see the photo of the prototype in **fig. 10**).

The equipping of the board should start with those components which are difficult to get at, or that must be soldered to ground on all sides, or that lie in the vicinity of the wall. It is preferable not to provide an IC holder for the S 042 P. The five pins which are shown connected to ground, are splayed out at right-angles on to the ground plane surface and soldered on to it.

Since the ferrite beads are always placed in the supply line to each individual stage, it is advantageous to include them one by one as the stages are tested.

Upon switching on the completed unit, the stabilized + 5 V and + 9 V supply rails are tested to ensure that they are functioning correctly. The phase loop is then opened by temporarily removing R6 from the PI controller's output and connecting it with the wiper of a test poten-

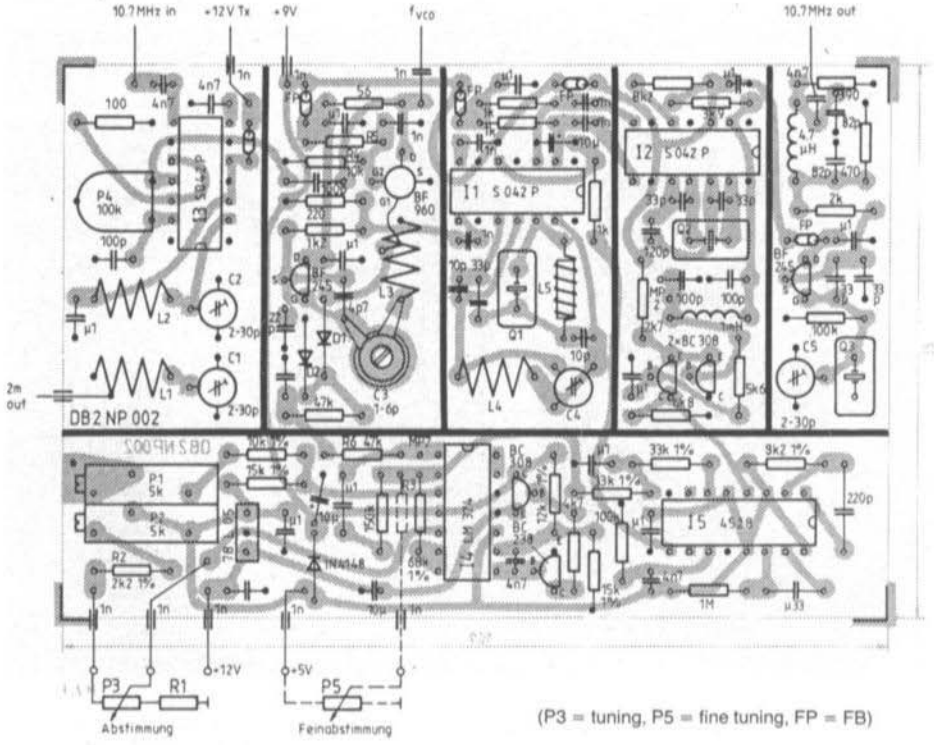


Fig. 9: The component layout for PCB DB 2 NP 002 for the circuit of fig. 8

tiometer, of about 10 to 50 kΩ, which lies across the + 9 V supply.

Now, the VCO output is connected to a frequency counter and both frequency and the voltage at the wiper of the test potentiometer monitored simultaneously. Capacitor C3 is so adjusted that at f_H = 133.8 MHz 7 V (approx.) is obtained. At f_L = 133.3 MHz, the reading should then be 3 V (approx.). If, owing to the tolerances of D1 and D2, this voltage range differs from 4 V, the tapping point of L3 should be shifted towards the cold end of the inductance to reduce the voltage range and towards the hot end to increase it.

Following this preliminary tuning, the VCO is set to 133.5 MHz (approx.). By means of the voltage divider R4/R5, the VCO output voltage can be varied within certain limits. A guide line for this voltage is that provided by the gate-2 potential of the BF960, i.e. 1.8 V which entails a 39 kΩ resistor for R5.

Now, L4 is loosely coupled by an inductive loop to a sensitive frequency counter. By carefully adjusting the capacitor C4, the 5th overtone of the crystal Q1 is excited. It will be immediately apparent from the behavior of the frequency counter if the oscillator is oscillating wildly or

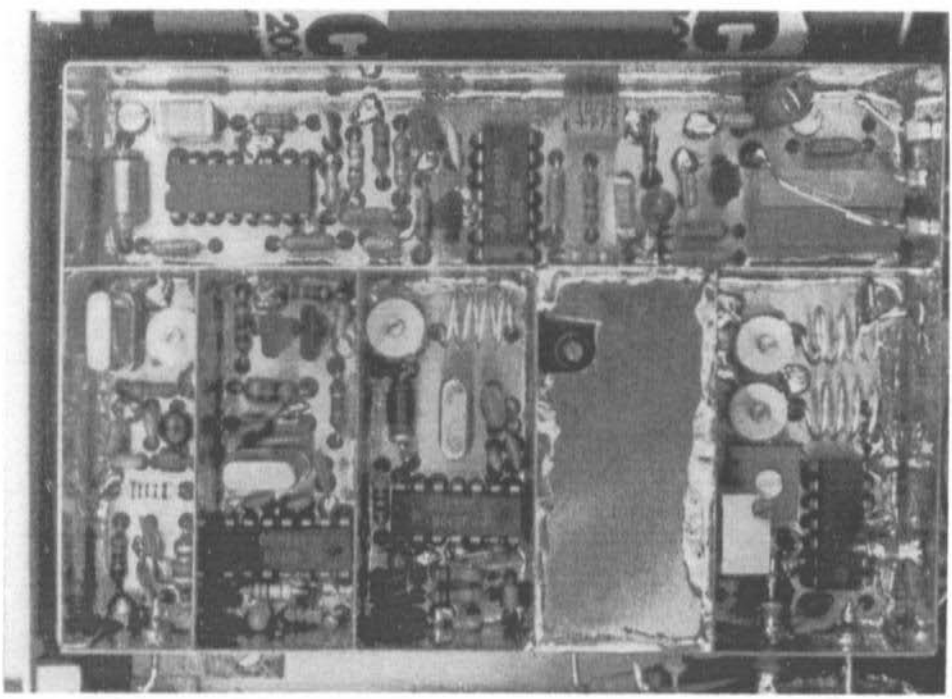


Fig. 10: Prototype of the fully equipped PCB DB 2 NP 002. The oscillator stage is largely hidden away under the soldered-on compartment cover. The trimmer C3 must be accessible. This is achieved by providing a 5 mm hole in the PCB and reversing the trimmer so that its slotted spindle protrudes through the ground-plane.It is then soldered to the track-side.

suddenly locks in to the crystal's overtone frequency. This stage is, in fact, somewhat critical. It may be necessary to match RFC L5, the crystal's capacity compensator, to the crystal in use. Suffice it to say that the stable operation of this stage is absolutely necessary for a stable VCO operation!

The second mixer/oscillator is, on the other hand, without any problems whatsoever. The down-converted VCO, tunable over 200 kHz, can be inspected with an oscilloscope at TP 1. Its amplitude should not be smaller than 0.2 V_{pp}. The amplitude can be maximized by a careful adjustment of C4 which makes the first mixer/oscillator oscillate at full power.

A slightly rounded square-wave signal should now be seen at pin 4 of the 4528 of 4 to 5 V_{pp} amplitude. This triggers the monostable. At its output (pin 6), clean square-wave signals should appear in synchronism with the trigger signal and whose period (1.5 μs) is set by R7 and C7.

The 12 V supply is now disconnected and the PLL loop closed by removing R6 from the test potentiometer and connecting it to the PI controller's output.

After re-applying the 12 V supply, there should be 3 to 7 V at TP 2. If it is only a few mV or nearly 9 V, i.e. at the extremes of the given range, it is an indication that the VCO, for some reason or other, is not phase-locked.

Now, the tuning for f_H = 133.8 MHz and f_L = 133.3 MHz can be carried out with P1 and P2 in exactly the same way as described earlier. At the band extremities it would appear as if the multi-turn pot'meter is non-linear. It may be therefore advisable to set this pot'meter (P3) to the first graduation on the scale and then with P1 set the frequency to 133.35 MHz. In the same manner, P3 is then set to scale graduation 9 and P2 is then adjusted for 133.75 MHz. The extreme frequencies then deviate but slightly from 133.3 and 133.8 MHz but the rest of the scale is much more precise. It should be noted, however, that potentiometers P1 and P2 are not so isolated from each other and this, of course, means more iteration from high to low scale until a satisfactory calibration has been achieved.

In conclusion, the stages which do not belong to the actual VCO are tuned. Capacitor C5 is tuned to obtain exactly 10.70000 MHz from the crystal oscillator Q3. If its output is then, provisionally, fed into the 10.7 MHz input of the transmit mixer, a signal at 144.0 to 144.5 MHz should appear at the mixer's output. Capacitors C1 and C2 should then be set to achieve maximum output power and P4 adjusted for the best possible carrier (133 MHz) suppression. The latter adjustment is carried out by withdrawing crystal Q3 and adjusting P4 for a minimum output. By increasing R8 it is possible to attenuate the 144 MHz output signal and thereby saving on battery supply requirements.

8. OPERATION AND CIRCUIT VARIANTS

The VCO as described was tried by the author in an SSB/AM home-constructed transceiver. After an initial warm-up period of about 3 minutes, it was not necessary to keep re-adjusting the frequency tuning. Components that were toleranced 1 % were, in fact, replaced by 0.1 % tolerance and a thermal drift of ± 50 ppm. A thermal compensation process such as that mentioned in chapter 2, is not, however, absolutely necessary for amateur applications. It would only become unavoidable if, during the course of operation, that large extremes of temperature were encountered and also a large store was laid upon the scale marking being absolutely correct at all times. Such temperature variations are only possible in practice, when the equipment is taken from a warm room into outside air with a temperature of say − 10° C in winter or a summer temperature of + 50° C and operations are immediately commenced. This sort of operational practice will certainly require corrective frequency procedures to allow drift-free operation following switch-over breaks.

A rule of thumb for the optimal dimensioning of the monostable's R7 and C6 to achieve good thermal stability (such as e.g.: 50 Ω NTC in series with R7) cannot be given, as the frequency drift

is dependent upon the inter-relationship between too many components. The correction must be determined individually for each VCO by a series of measurements in a thermally controlled enclosure.

The 10-turn potentiometer has been specified with a ± 0.25 % linearity. As the VCO tuning range covers 500 kHz, an error uncertainty of this magnitude can be expected in the frequency setting due to this cause alone i.e. ± 1.25 kHz. This can be only be circumvented by an individual scale calibration or by a correction scale/table. The author's prototype exhibited a scale error of ± 1 to 2 kHz, the precise error depending upon the frequency setting within the 500 kHz range. A more exact frequency setting could be obtained by the employment of a ± 1 % linearity potentiometer or with a digital frequency read-out. As the VCO signal is downwards translated in any case, only a very slow frequency-counter chip is required, e.g. the 5-digit LCD module EA 6077 which is able to work up to 4 MHz and takes a supply current of only 2 mA (2). To bring it into operation requires only the provision of a buffer stage, e.g. a CMOS inverter stage connected with the monostable's trigger signal. To enable the frequency to be read easily, a specially selected combination of Q1 and Q2 is necessary in order that the exact band edge frequencies, e.g. 100.0...600.0 kHz or better 0...500.0 kHz are produced. The latter then would correspond to a working frequency of 144.0...144.5 MHz.

As the VCO is very easily frequency modulated (see part 1, fig. 1), it would appear a natural reaction to build a circuit variation which will cover the full 2-metre band for use in an FM-transceiver. The required modifications for this application are to be seen in **fig. 11**. The first mixer oscillator is crystal-stabilized at a frequency of 144 MHz minus the IF, i.e. in this case at 133.3 MHz supplied, if necessary, from an external XO. The second mixer oscillator oscillates with the IF, when in the TX mode, i.e. 10.7 MHz. If now, the monostable's time constants are reduced to 400 ns and the limit frequency of the low-pass filter following the 2nd mixer increased to 2.5 MHz, the VCO is able to be tuned to cover a frequency range from 144 to 146 MHz.

The provision of a frequency offset for radio transponder operation is effected by a 10.1 MHz (i.e. 10.7 MHz minus 600 kHz) crystal. In the receive mode (RX), the second mixer oscillator is totally unbalanced and the oscillator part of the circuit suppressed in order that the mixer input signal passes spectrally unaltered to the mixer output. The consequence of this is that the VCO supplies the frequency range 133.3 MHz to 135.3 MHz, i.e. the working frequency minus the IF, to the receive mixer.

As the VCO deviation is multiplied by four times in the process of arriving in the vicinity of the band to be used, the tapping point for the varicap diodes D1 and D2 on the inductor L3 must be made higher up the coil. In addition, a second pair of varicap diodes has the job of providing a large frequency shift, by an amount equal to the IF, during TX/RX switching in order that the PI controller can retain control, within its output voltage limits, over the loop frequency for both TX and RX. Due to the shifting of the VCO spectrum by an amount equal to the IF, an up-conversion mixer for TX is rendered superfluous and therefore obviating the spurious signal problems attending mixer stages.

In SSB equipment, this upward frequency translation is unavoidable as the SSB processing takes place invariably at the IF level. As the monostable trigger signal lies, in both the case of "send" and "receive", in the range 0 to 2 MHz, it is easily retro-fitted with the LCD counter module EA 6077. The working frequency is arrived at by the addition of 144 MHz to the displayed frequency. As the tuning range now is 0 to 2 MHz and not as previously considered, 0 to 500 kHz, the zero is not very exact and could read out at several kHz. The other case entails the coupling and de-coupling capacitors having an excessively high value thus making the VCO unstable in its lower tuning range. (There is also the possibility of a random transition to the other sideband: X-coupled PLL). When tuning the VCO for the indicator translation 0 to ... MHz, the trigger signal lower frequency limit should commence at about 5 kHz for the SSB version and about 25 kHz for the FM version. The lowest working frequency would then be 144.005 MHz (SSB) or 144.025 MHz (FM) and not 144.000 MHz.

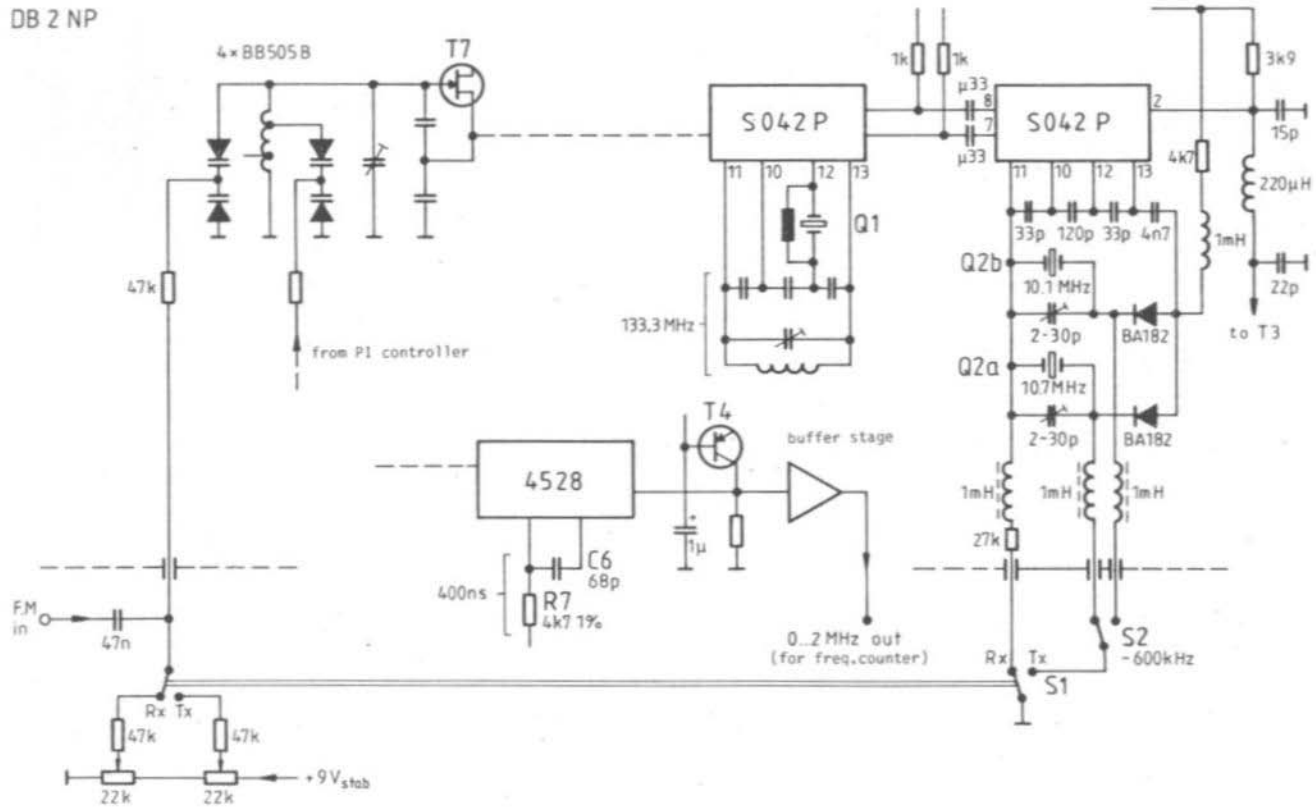


Fig. 11: Modification of the VCO of fig. 8 for an 144 - 146 MHz FM-transceiver having a 0 - 2 MHz output for a frequency counter. Switch S1 is the IF shift for the RX/TX change-over. Switch S2 enables the 600 kHz off-set required on TX for relay operation. The 9 MHz IF version requires Q 2a and Q 2b to be changed to 9 MHz or 8.4 MHz. The up-conversion mixer (IC3) and the carrier oscillator stage (T1) shown in fig. 8 are redundant.

9. COMPONENTS

I1...3:	S 042 P (Siemens)
I4:	LM 324 (various manufacturers)
I5:	4528 (e.g. Fairchild, Valvo)
I6:	78L05 (various manufacturers)
T1, T7:	BF 245
T2:	BF 960
T3...T5:	BC 308 or pnp equiv.
T6:	BC 238 or npn equiv.
D1, D2:	BB 505 B (approx. 2...11 pF)
D3:	1N 4148 or Si. equiv.
C1, C2, C4, C5:	Trimmer 2...30 pF, red, RM 5 mm (Valvo)
C3:	Cer. Trimmer 1...6 pF, multi-turn
C6:	low-loss, styroflex, polypropolane etc.
L1:	4 turns tapped 1.5 turns from cold end
L2:	4 turns
L3:	4 turns tapped 1 turn from cold end
L4:	4 turns

Note: Inductors L1...L4 are to be wound with 0.8 mm silvered wire, L3 ist to be wound on a 6 mm former with 1 mm silvered wire

L5:	10 turns Cu.lacq. 0.3 mm wire wound on 27 kΩ resistor of 3 mm dia.
L6, L7:	Fixed inductor with ferrite core (e.g. Siemens)
FB:	Ferrite bead
Q1:	24.576 MHz fundamental crystal
Q2:	10.24 or 10.245 MHz fundamental crystal
Q3:	10.7 MHz fundamental crystal

Note: For a 9 MHz IF, Q3 should be 9 MHz and Q2 should be either 11.94 or 12.0 MHz

P1, P2:	Preset pot'meter RM 5/7.5/ 2.5 mm
P3:	10-turn potentiometer with panel scale 0...10
P4:	Preset pot'meter, carbon film, RM 10/5 mm (Piher)

Note: All resistors marked 1 % are to be metal-film with TK = 100 ppm/°C, or better 0.1 % with TK = 50 ppm/°C.

All feed-thro' capacitors are approx. 1 nF, solderable.

The 4 feed-thro's are PTFE.

The tin-plate box is 74 x 111 x 30 mm with two covers and is fitted with a compartment divider wall 20 x 109 mm and 4 sub-dividers 20 x 47 mm. An internal cover 22 x 47 mm should also be provided for the VCO stage.

10. REFERENCES TO PART 2

(1) Oppelt, Ralph, DB 2 NP:
The Generation and Demodulation of SSB Signals using the Phasing Method.
VHF COMMUNICATIONS 19 (1987),
Edition 2, Pages 66 - 72
Edition 3, Pages 130 - 140

(2) (not mentioning the author)
4-MHz-Frequenzzähler
Elektor, Edition 6/88, P. 20 - 21

EUROCARD

We accept **VISA Credit Card, Eurocard** (Access/Master Card) and only require the order against your signature, card number and its expiry date.

VHF COMMUNICATIONS / UKW-BERICHTE

VISA